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A SYNCHRONOUS DETECTION SYSTEM 12T

FOR HIGH FREQUENCY USE

A THESIS

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Loyce Archibald Turner

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A SYNCHRONOUS DETECTION SYSTEM

FOR HIGH FREQUENCY USE

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ABSTRACT

The purpose of this thesis is to design and test a synchronous detection system which will operate linearly over a range of 80 db.

In research and development work it is frequently necessary to record a signal whose amplitude may vary over wide ranges. This is usually done by modulating the signal at an audio frequency and using a crystal diode, bolometer, or other square-law device as a detector and using the output from this detector to operate a logarithmic recorder. The dynamic range of most commercial logarithmic recorders is about 80 db, and, when used with a square-law detector, the range of signal level that can be recorded is only about 40 db. This same recorder, when used with a synchronous detector, has a dynamic range of 80 db.

The synchronous detector requires a constant amplitude reference signal of the same frequency and phase as the signal being detected. A separate path from the transmitter to the detector is usually necessary, although an oscillator which is synchronized with the received signal may be used instead. The system described in this thesis makes use of two channels. One of the channels consists of a high-gain amplifier and a limiter and provides the constant amplitude reference signal for the synchronous detector. The other channel contains a

modulator in which the signal is chopped at a frequency of 2,500 cycles. The outputs of the two channels are combined in the synchronous detector. The output of the synchronous detector is a 2,500-cycle signal which may be used to operate a logarithmic recorder.

This synchronous detection system was designed for use as the intermediate frequency amplifier of a microwave receiver. It operates at a frequency of 36.5 megacycles and has a bandwidth of 4.0 megacycles. The system also includes an automatic frequency control unit which provides sufficient output voltage to control a reflex klystron oscillator.

This synchronous detection system gives very good results over the 80 db range between -5 dbm and -85 dbm. Its departure from linearity over this range is about 1.0 db at 36.5 megacycles and about 4.0 db near the band edge.

CHAPTER I

INTRODUCTION

The problem of obtaining linear detection.--Most of the common systems for detection of a high frequency signal depend upon rectification or the use of a non-linear device such as a vacuum tube operating near cut-off. These systems, and also power sensitive devices such as bolometers and thermistors, tend to have a square law response at low signal levels. With a square law detector, it is very difficult to detect a signal below about 0.001 microwatt. The square law detector is most useful at power levels greater than 0.01 microwatt.

If a detector gave a linear rather than a square law response, it should be possible to detect a signal at a much lower power level. This paper deals with the synchronous detector as a means for obtaining this linear response. It is one of the few types of detectors that has a linear response over a wide range.

The synchronous detector.--A synchronous detector is a mixer circuit in which the oscillator and signal frequencies are the same, and their relative phase is held at either zero or one hundred and eighty degrees. It could also be considered as a detector producing rectification. The synchronous detector described in this thesis uses two 6AS6 pentode tubes. A variable transconductance is obtained by applying a variable voltage to the suppressor grid. This voltage varies the division of the cathode current between the screen and the plate while having negligible effect on the value of the cathode current.

When a square-wave voltage of sufficient amplitude is applied to the suppressor grid of a 6AS6 tube, a square-wave plate current will flow with a minimum value of zero. Under these conditions, with a sinusoidal voltage applied to the grid, and with a relative phase angle of ϕ degrees with respect to the square-wave, the average plate current of the tube could be expressed as

$$\omega/2\pi \int_{\phi}^{\pi+\phi} A \sin \omega t \, dt = A/\pi (\cos \phi) + K$$

Under these conditions it can be seen that if the relative phase angle can be held constant, an exact reproduction of any modulation, A, on the input signal will be obtained. K is the static plate current of the tube with no signal applied to the control grid.

A more complete analysis, using a series expansion of the reference signal, is given in the next chapter.

If the reference signal applied to the suppressor grid is sinusoidal and its amplitude is such that the suppressor grid-to-plate transconductance is approximately constant, the control grid-to-plate transconductance will be modulated by the suppressor grid voltage and may be expressed in the form

$$g_m = B + CD \sin \omega t$$

where B and C are constants and D is the amplitude of the suppressor grid voltage. If the control grid voltage is

$$E \sin (\omega t + \phi)$$

the plate current may be expressed as follows:

$$i_p = [E \sin (\omega t + \phi)][B + CD \sin \omega t]$$

$$i_p = BE \sin (\omega t + \phi) + (CDE \sin \omega t)(\sin \omega t + \phi)$$

$$i_p = BE \sin (\omega t + \phi) + \tfrac{1}{2}CDE \cos \phi$$

$$- \tfrac{1}{2}CDE \cos (2 \omega t + \phi)$$

If the output current from the tube is passed through a low-pass filter, the output current from the filter may be expressed as

$$i = \tfrac{1}{2}CDE \cos \phi$$

The output is proportional to the product of the amplitudes of the oscillator and the signal and the cosine of their relative phase angle. If the oscillator amplitude and relative phase angle can be held constant, the output will be directly proportional to the input. Usually when a synchronous detector is used a separate path is provided from the oscillator to the detector for the oscillator or reference signal. Slight variation in path length may cause errors in the relative phase shift at the detector and result in non-linear response. In the unit to be described, this difficulty was overcome by using a single path from the transmitter to the receiver and dividing it within the receiver.

Block diagram of a synchronous detection superheterodyne.-- In the block diagram of the receiver, Fig. 1, an unmodulated high frequency signal comes from the antenna to a mixer as in a conventional superheterodyne receiver. Automatic frequency control is included to simplify microwave operation when a klystron local oscillator is being used. The signal from the mixer is fed into two channels. One is used to produce the oscillator or reference signal for the detector. It is a synchronous tuned amplifier using seven 6AK5 tubes. The center frequency is 36.5 megacycles. The first stage is a cascode amplifier using a triode connected 6AK5 to obtain a low noise level. The last two stages are designed to act as limiters and give a constant output.

Loading resistors are paralleled with all tuned circuits to produce a bandwidth of about four megacycles. The gain is about 130 db. The output of the reference channel is fed into both the synchronous detector and the automatic frequency control circuit.

The automatic frequency control circuit consists of a ratio detector followed by a direct current amplifier. The output voltage swing is approximately 40 volts peak to peak, which is sufficient for most common reflex klystron tubes.

The other channel feeds into a modulator which chops the signal at 2500 cycles per second. This is equivalent to

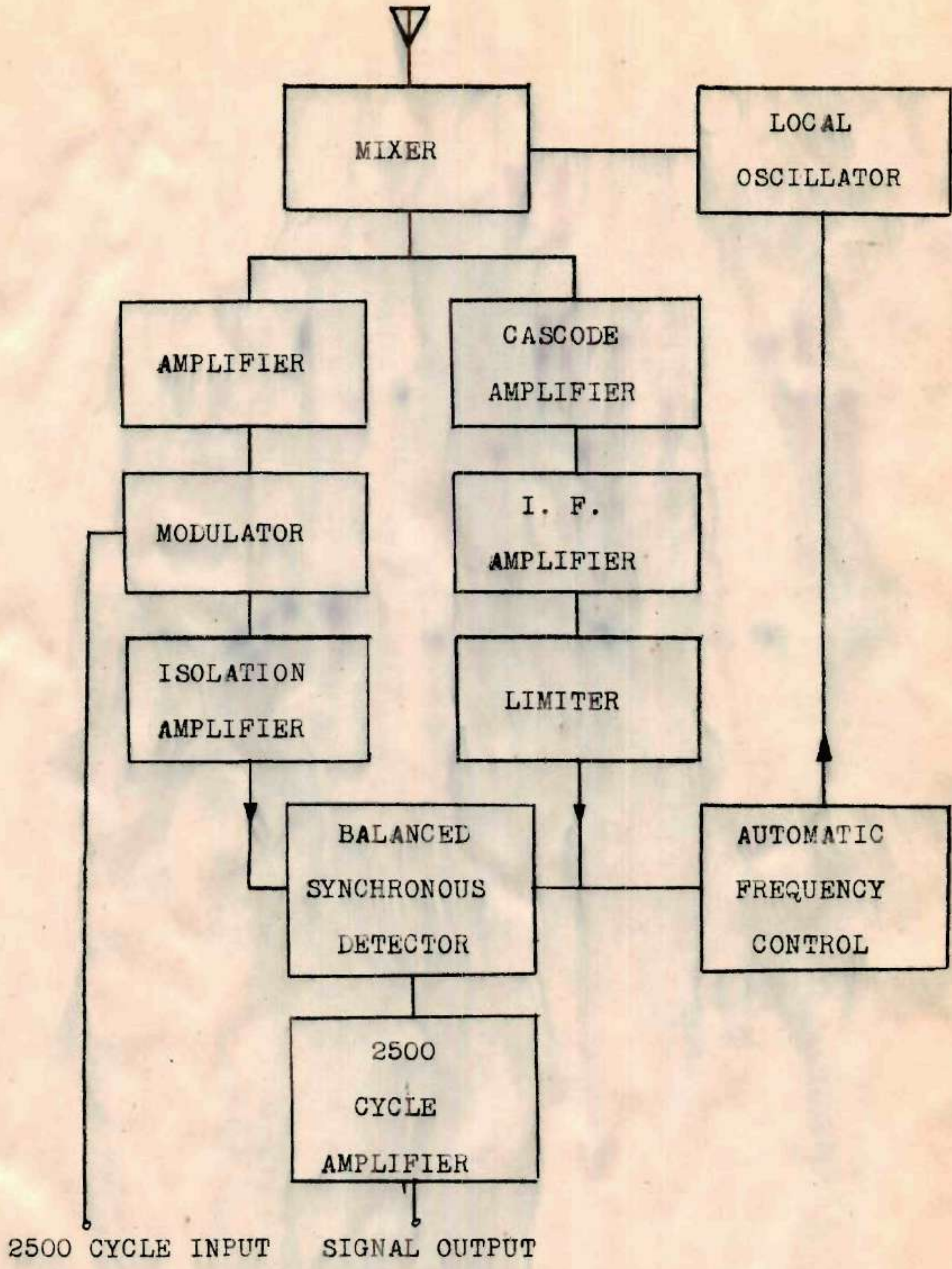


Fig. 1. Block Diagram of a Synchronous Detection Superheterodyne

square wave modulation at that frequency. The output of the modulator feeds into an isolation amplifier. This channel contains a phase shifting circuit which has the same phase shift versus frequency characteristic as the reference channel. This channel gives a push-pull output which is fed into the control grids of two 6AS6 tubes.

The two 6AS6 tubes are used as a synchronous detector. The signal to be measured is fed into the control grids of the two tubes. The output of the reference channel is fed into the suppressor grids in parallel. Push-pull output is taken from the plates. This arrangement tends to cancel noise that may be produced by rectification of the reference signal on the suppressor grid. The output of the synchronous detector is amplified by a tuned logarithmic amplifier and then recorded if it is desired to plot a very wide range of values.

Previous work with the synchronous detector.--During 1953, M. E. Brodwin, C. M. Johnson and W. M. Waters (1) of John Hopkins University, in an article on synchronous detectors, reported sensitivities of -110 dbm, about 30 db better than a crystal detector. Linear response over a range of 60 db was easily attained. The system used required two signals to the receiver from the transmitter. One was modulated while the other was carefully controlled in amplitude and phase shift. Tests were conducted at frequencies up to 99 kilomegacycles.

Earlier work on the synchronous detector was done by

D. G. Tucker (2) during 1952. He used a system very similar to that used by Brodwin, Johnson, and Waters. He claims better signal to noise ratio on pulsed signals than was obtained with conventional detectors.

In 1951, R. A. Smith (3) published an article on coherent and incoherent detectors, in which he said he obtained a considerable improvement in signal-to-noise ratio at signal-to-noise ratios of less than unity when using a coherent detector.

During 1949, W. C. Michels and E. D. Redding (4) published an article on an improved synchronous detector. It had a very narrow bandwidth and operated at 800 cycles per second.

The original work on synchronous detection was done by W. C. Michels and N. L. Curtis (5) in 1941. A low frequency was used then as in their later work.

The unit to be discussed in this paper will differ from previous detectors of this type in being wide band and requiring a single signal from the transmitter. It is intended to be the intermediate frequency amplifier of a microwave superheterodyne system, but this would not be necessary if it were used at frequencies below about fifty megacycles.

CHAPTER II

ANALYSIS OF CIRCUITS

The reference channel.--The reference channel takes the place of the separate reference path from the transmitter, which is usually required by a synchronous detector. It contains a high gain synchronous-tuned amplifier. A cascode amplifier is used as the input stage of the reference channel amplifier because it has a very low noise output. It gives considerably better sensitivity than would be obtained if a pentode were used. The cascode stage is followed by five stages of amplification in which pentodes are used. The last two of these stages are limiters and have a constant output of about 15 volts over the dynamic range of this unit. The output of this amplifier is the reference signal which is used by the synchronous detector. Fig. 2 shows the circuit diagram of the reference channel amplifier.

A complete discussion of the cascode amplifier may be found in either Vacuum Tube Amplifiers by Henry Wallman and George E. Valley, Jr. (6) or "A Low-Noise Amplifier" by Henry Wallman, Allan B. Macnee and C. P. Gadsden (7), in the June, 1948, issue of the Proceedings of the Institute of Radio Engineers. In the circuit used in this unit, the type 6AK5 tube is used as a triode with the screen tied to the plate.

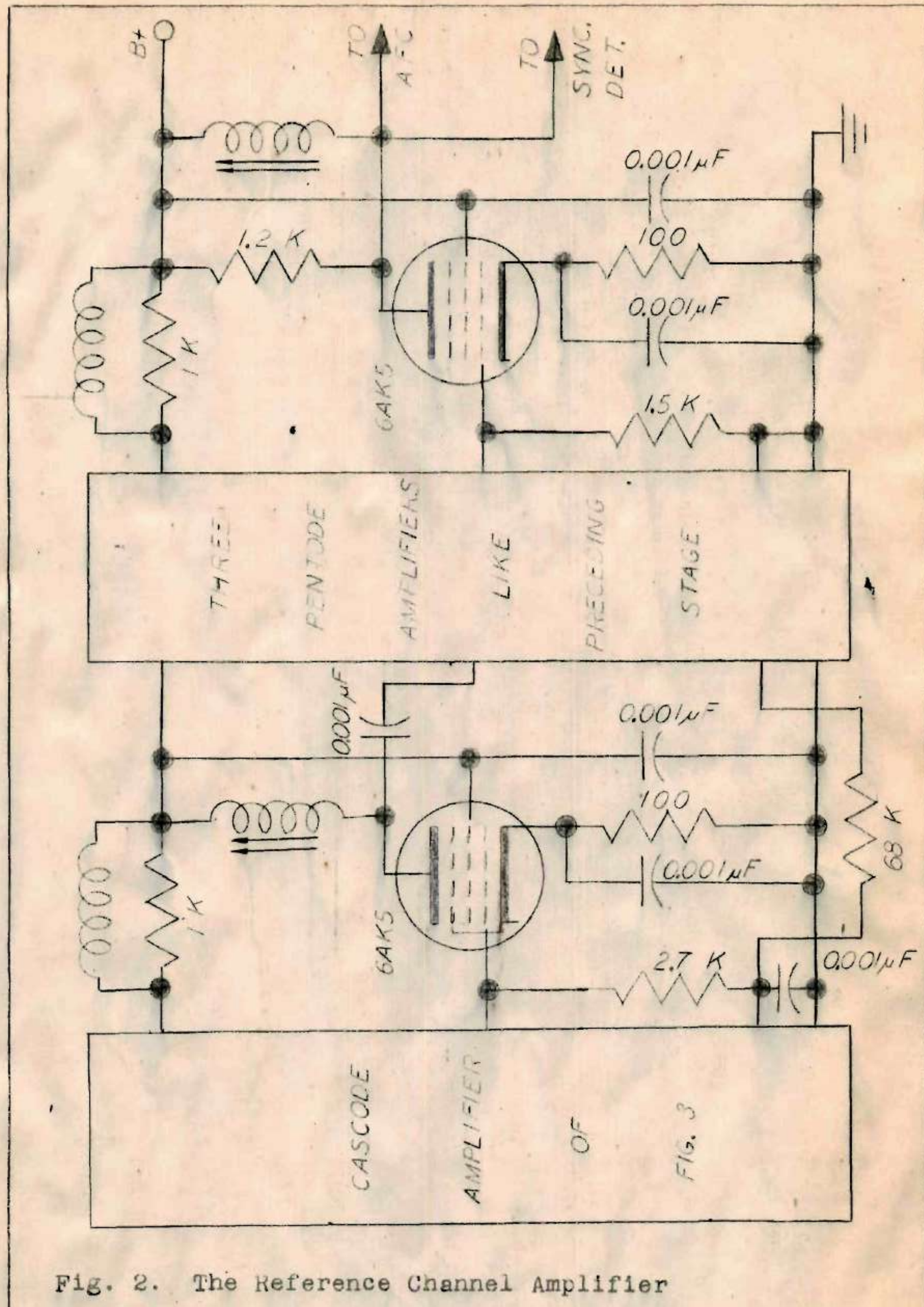


Fig. 2. The Reference Channel Amplifier

The 6AK5 triode has an equivalent noise resistance of 385 ohms. When it is operated as a pentode the equivalent noise resistance rises to about 1880 ohms, and the maximum sensitivity of an amplifier using the tube would be only forty-five per cent of that that could be obtained if the tube were operated as a triode. The second tube of the cascode is a type 6AS6 which is used as a triode by tying the screen grid, suppressor grid and plate together. The amplifier is tuned to 36.5 megacycles by variable inductance. The circuit diagram of the cascode amplifier used in this unit is shown in Fig. 3.

The operation of the cascode amplifier is very similar to that of a pentode tube which has a transconductance and equivalent noise resistance equal to those of the input tube. The second tube of the cascode has very little effect on either the noise factor or the gain of the circuit. The input impedance of the grounded-grid or second tube is approximately the reciprocal of its transconductance, if this tube has a load resistance which is low in comparison with its plate resistance. This input impedance serves as the plate load of the first tube and is very much lower than its plate resistance. The transconductance of the input tube is g_{m1} and its load resistance is R_{L1}. The transconductance of the second tube is g_{m2} and its load resistance is R_{L2}. The gain of either stage may be expressed as

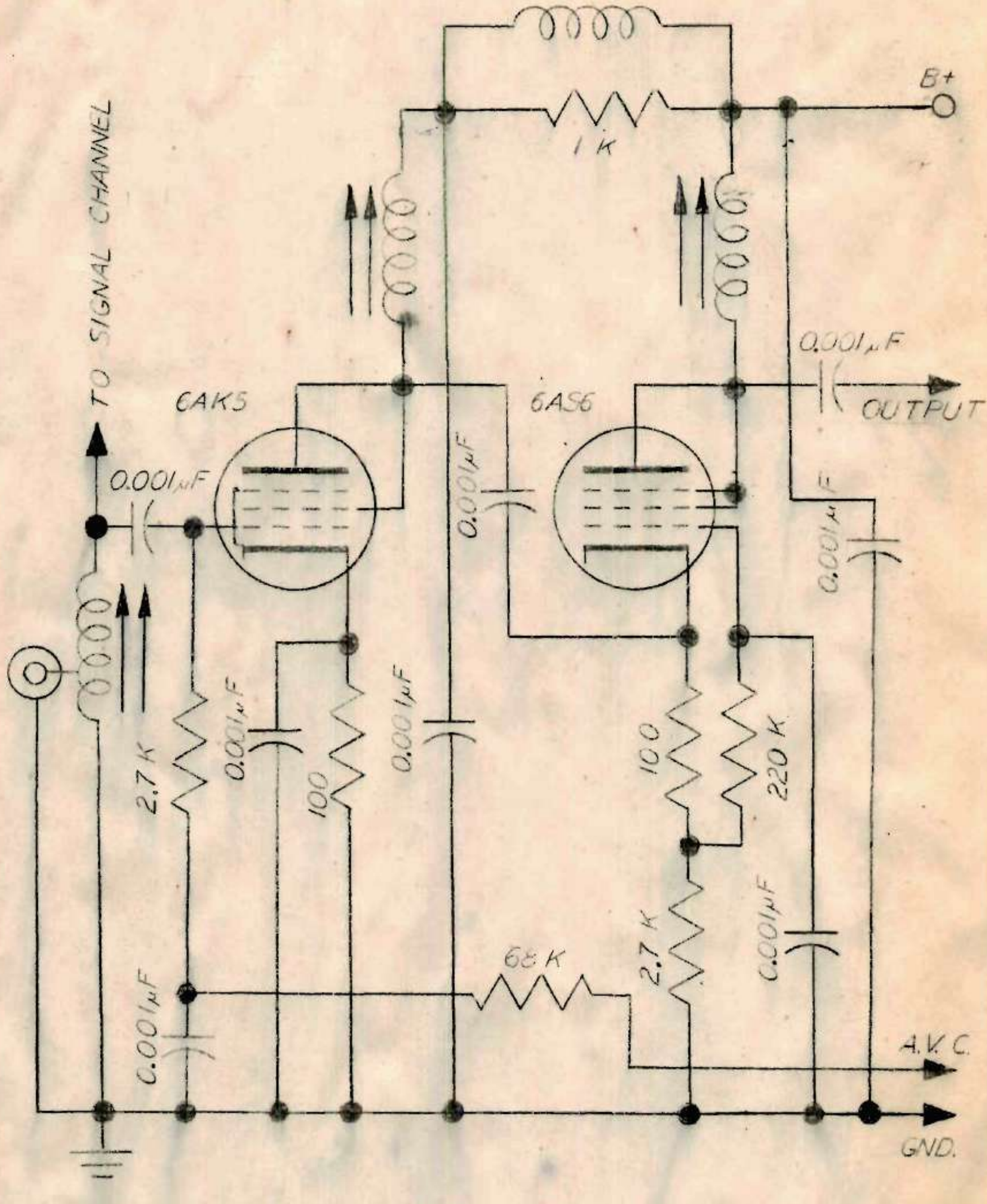


Fig. 3. The Cascode Amplifier

$$A = g_m R_L$$

The gain of the first tube is

$$A_1 = g_{m1} R_{L1} = g_{m1}/g_{m2}$$

The gain of the second tube is

$$A_2 = g_{m2} R_{L2}$$

The gain of the cascode amplifier is

$$A_1 A_2 = g_{m1} R_{L2}$$

In the circuit used in this unit, g_{m1} has a value of 0.0067 mho and R_{L2} is 2700 ohms, and the voltage gain of this stage is

$$A_1 A_2 = (0.0067)(2700)$$

$$A_1 A_2 = 18.1 \text{ or } 25.1 \text{ db}$$

It should be noticed that the plate of the first stage and cathode of the second stage are tuned to 36.5 megacycles. The input impedance of the grounded-grid amplifier is about the same magnitude as the reactance of the stray capacity in this circuit. The Q of this circuit is approximately one since

$$Q = R\omega C$$

The remaining five stages of the reference channel amplifier use 6AK5 pentodes. They bring the total gain of the reference channel amplifier up to 130 db.

There are seven tuned circuits in this channel that introduce phase shift if the frequency is shifted to either side of the center frequency. In order that it may be compensated for, the phase shift-versus-frequency characteristic of this amplifier will now be calculated.

The pentode tubes have very high plate resistances and may be considered as constant current sources. The phase shift of the reference amplifier will therefore be equal to the sum of the phase angles of the impedances of the tuned circuits used as loads in this amplifier.

Consider the tuned circuit as a parallel inductance, capacitance, and resistance. At the resonant frequency in radians per second, ω_o, the reactances of the inductance and the capacitance are equal. The resistance is R. The applied frequency is ω.

The admittance, Y, of the tuned circuit may be written

$$Y = 1/R + j\omega C - j/\omega L$$

If the resonant frequency is f_o and the applied frequency is f, this equation may be written as

$$Y = 1/R + (j\omega_o C)(f/f_o - f_o/f)$$

This equation may be simplified by using the definition

$$\Delta f = f - f_o$$

If Δf is much smaller than f_o, the admittance of the tuned circuit may be expressed as

$$Y = 1/R + j2\omega_o C\Delta f/f_o$$

The phase shift, θ, of the parallel resonant circuit may be expressed as

$$\theta = -\tan^{-1} B/G$$

where B is the imaginary part and G the real part of Y. If the values of B and G are substituted, the phase shift may be written

$$\theta = -\tan^{-1} 2\omega_o CR\Delta f/f_o$$

The Q of a parallel resonant circuit at its resonant frequency is defined as

$$Q = R\omega_o C$$

If Δf is expressed in megacycles and the value of f_o, 36.5 megacycles, used in this unit is substituted in the expression for phase shift, θ becomes

$$\theta = -\tan^{-1} Q\Delta f/18.25$$

For small values of $Q\Delta f$, this expression may be replaced by

$$\theta = -3.15\, Q \Delta f$$

where θ is the phase shift in degrees. From the above equation, and the Q of unity for the tuned circuit in the cathode of the grounded-grid stage, the phase shift of the first tube in the cascode amplifier can be calculated to be 3.15 degrees per megacycle. The plate circuit of the second tube has a Q of 5, which results in a total phase shift of 18 degrees per megacycle for this stage.

By the above method, the phase shift of the five remaining tuned circuits may be determined. Three stages have a phase shift of 15 degrees per megacycle and two stages have a phase shift of 9 degrees per megacycle. The total phase shift for the entire reference channel is approximately 81 degrees per megacycle. This approximate result is useful, since the tolerances in stray capacitance and circuit components would cause the phase shift to differ slightly from the calculated value. A difference of 10 degrees in the phase shifts of the reference channel and signal channel would decrease the output voltage about 1.5 per cent.

The overall gain of the reference channel, including the voltage step-up in the input circuit, is about 130 db. This gain, with the limiters, gives very nearly constant output over an input range of 80 db.

The output of the limiters is applied to the suppressor grids of the synchronous detector and to an automatic frequency control circuit.

The automatic frequency control uses a 6AK5 amplifier to drive a ratio detector. A complete discussion of the ratio detector may be obtained from either _Frequency Modulation_ by Nathan Marchand (8) or _F-M Transmission and Reception_ by John F. Rider and Seymour D. Uslam (9). The circuit diagram is shown in Fig. 4.

The ratio detector uses two 1N48 germanium diodes as rectifiers. They are followed by a two-stage direct-current amplifier which uses a 6J6 and a 12AX7 tube. The automatic frequency control chassis, including the secondary winding of the ratio detector transformer, the diodes, and the direct-current amplifier are designed to be operated at up to 4,000 volts to ground. This high-voltage insulation makes it possible to connect the output of this unit in series with the repeller voltage supply for a klystron oscillator. The klystron tube is the local oscillator of an external microwave converter and the synchronous detector described in this paper is used as the intermediate frequency amplifier and detector.

The ratio detector used in this unit is conventional with the exception that it has a center-tapped load resistor with the center tap grounded. Push-pull output is obtained

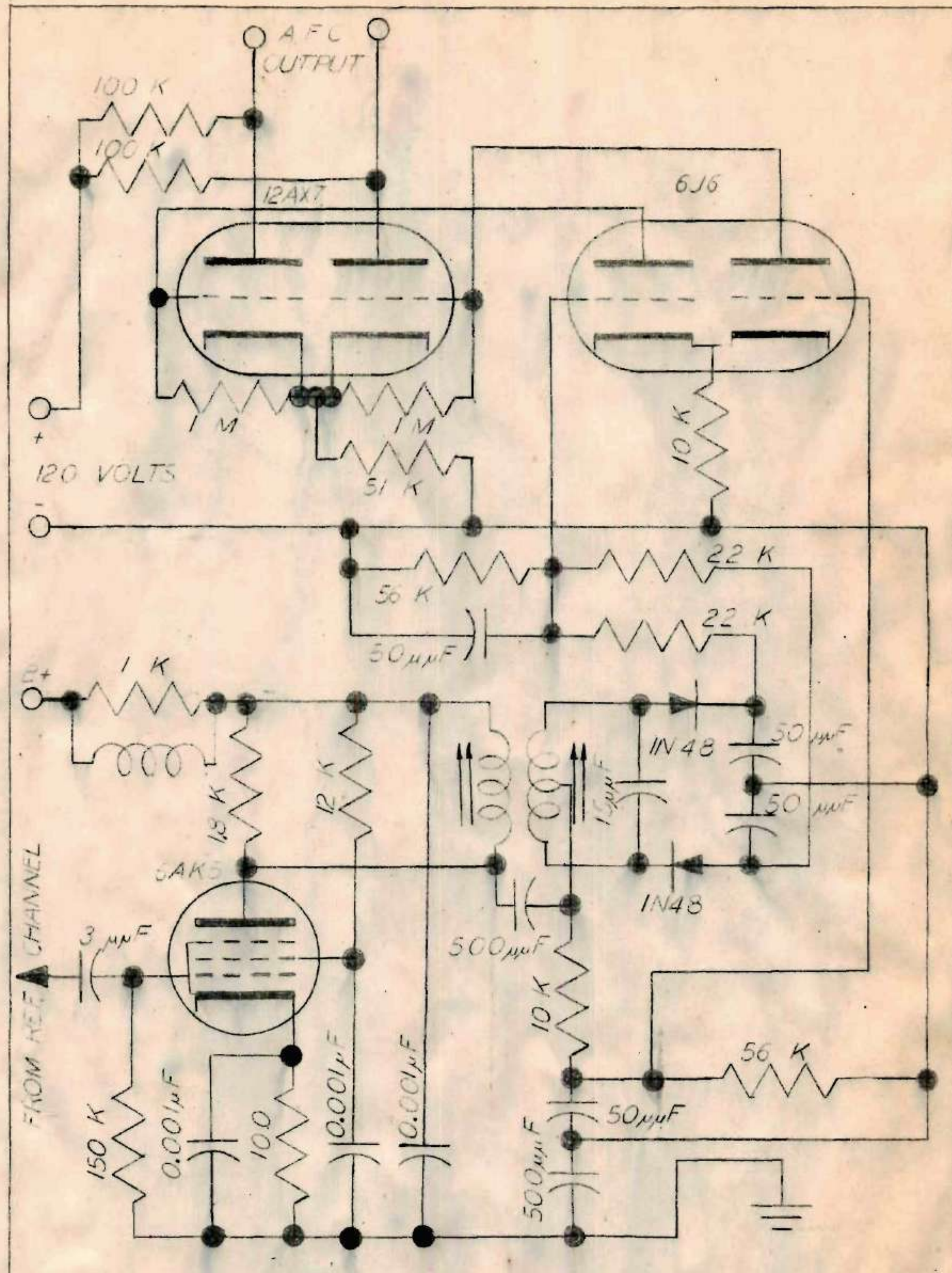


Fig. 4. The Ratio Detector and Automatic Frequency Control Amplifier

from the two ends of the load resistor.

The push-pull output from the ratio detector drives a 6J6 amplifier. The 6J6 amplifier has high gain, but overloads easily. Its output is directly coupled to the grids of a 12AX7 amplifier. The cathodes of the 12AX7 are biased 20 volts positive to provide plate voltage for the 6J6. The 12AX7 amplifier provides the automatic frequency control output, which is the difference of the voltages at the plates of the 12AX7 tube. A maximum correction of approximately 40 volts is obtainable. This output voltage is reached when the error in frequency is one megacycle and does not decrease until the frequency has changed by four megacycles. The automatic frequency control circuit is designed to maintain the incoming signal frequency between 36 and 37 megacycles for normal operating conditions. The bandwidth of the unit is much greater than this, but it is usually desirable to keep the error less than the 3 db that occurs at the band edge. The output from the automatic frequency control unit is connected to an external microwave converter by means of a shielded cable.

The operation of the unit may be shown by the following calculated example.

The output of the automatic frequency control amplifier is added to the repeller voltage of a reflex klystron tube which is used as the local oscillator of a microwave converter. If either this local oscillator or the signal being received tends to drift in frequency, the repeller voltage of the klystron tube will vary in such a manner that the output of the mixer will remain very close to the center frequency of the synchronous detector. Assume that a signal being received on a frequency of 10,000 megacycles drifts 15 megacycles. If a 2K25 klystron tube is being used as the local oscillator and the frequency of this tube changes one megacycle per volt change in repeller voltage, the drift at the intermediate frequency may be calculated from the response of the automatic frequency control unit. This response is about 160 volts per megacycle, so the drift at the intermediate frequency is

$$15/160 = 0.094 \text{ megacycle}$$

The change in the intermediate frequency for the above example is only slightly more than one per cent of the change that had occurred in the frequency of the carrier being received.

<u>The signal channel</u>.--The signal channel consists of an amplifier feeding into a delay line, a modulator, and an isolation amplifier. The circuit diagram is shown in Fig. 5.

This channel uses a 6AK5 amplifier in its input stage. The plate of this tube is connected to a transmission line, which has a phase shift-versus-frequency characteristic which

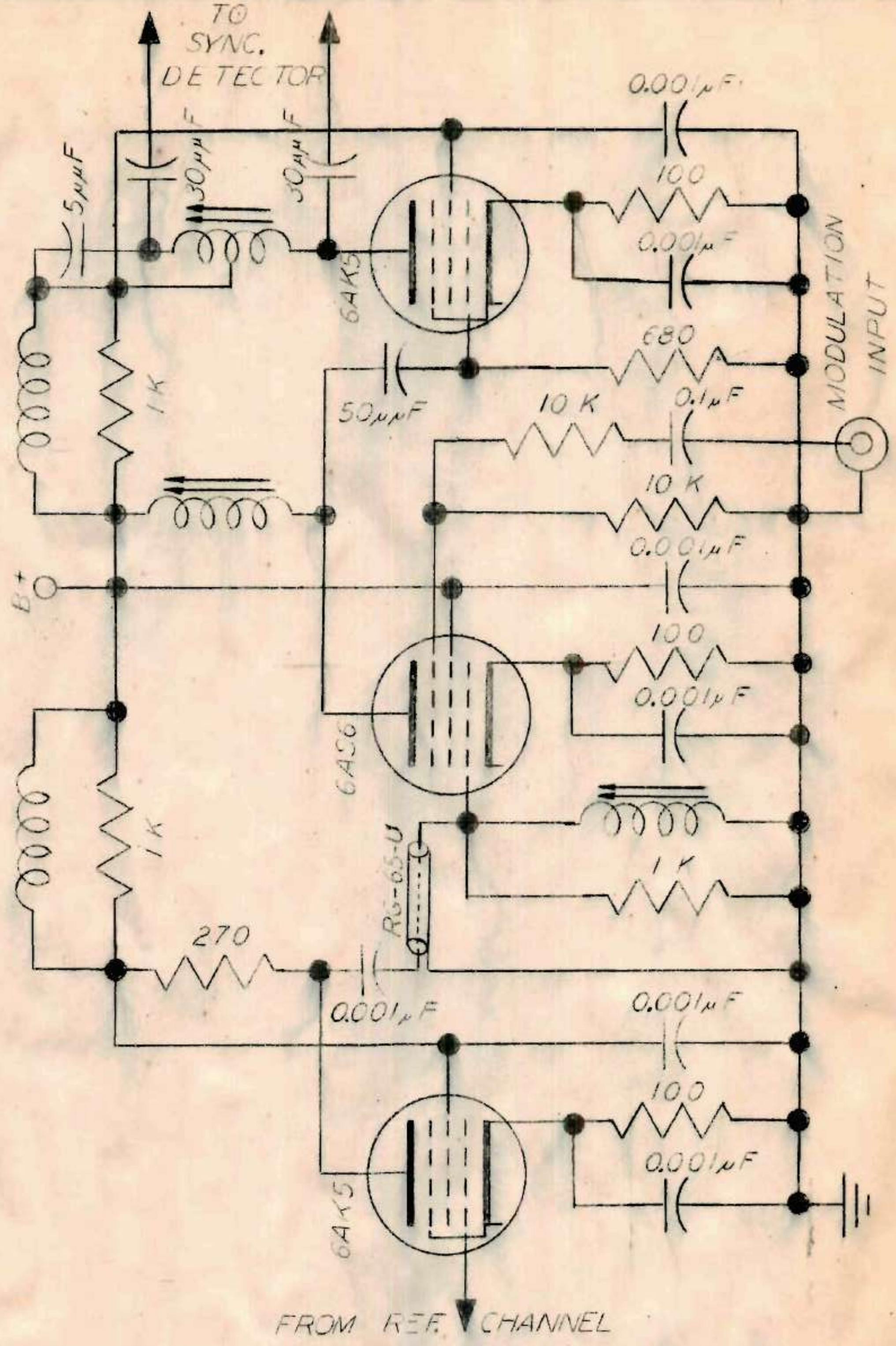


Fig. 5. The Signal Channel Amplifier

is close to that of the reference channel. It was shown in the first section of this chapter that the reference channel had a phase shift of 81 degrees per megacycle. This is equivalent to a transmission line of sufficient length to produce a phase shift of 0.225 wavelength when the frequency is changed 2.74 per cent. From these requirements the line length, X, can be calculated as

$$X = (X + 0.225)/(1.0274)$$

which gives

$$X = 8.2 \text{ wavelengths}$$

This corresponds to a time delay of 0.225 microseconds. RG-65-U coaxial cable is used as the delay line. It has a time delay of 0.042 microseconds per foot and a characteristic impedance of 950 ohms. For a delay of 0.225 microseconds, the length of cable required is 5.35 feet. This cable has an attenuation of 40 db per 100 feet at 36 megacycles. The delay line used in this unit has an attenuation of about 2.4 db. The cable is terminated with a resistive load of about 960 ohms, obtained by paralleling a 1000-ohm resistor with a tuned circuit.

Other systems could have been used to obtain the necessary phase shift, but the one which is used is probably the simplest. With a system in which the phase shift is linearly related to frequency, the total phase shift must be

2950 degrees. A lumped constant line made up of series resistance and parallel capicitance would either have too much attenuation or would require so many sections that it would be impractical. An arrangement using six tuned circuits could be used. This method would probably be the most satisfactory solution if the frequency used was below about ten megacycles. At frequencies near 36 megacycles, considerable difficulty was encountered in obtaining sufficient isolation between the signal channel and the reference channel. This is a result of the very great difference in the signal level of the reference channel and the signal channel, as will be discussed later. The addition of the six tuned circuits would make it very difficult to isolate the two channels. For this reason the use of the coaxial cable was considered the best solution.

The output of the coaxial cable feeds a type 6AS6 pentode, used as a modulator. The signal is chopped at a 2.5-kilocycle rate by a square-wave voltage applied to the suppressor grid. The plate circuit of the modulator is tuned to 36.5 megacycles by means of a 2.2-microhenry coil. A 50-micromicrofarad coupling capacitor from the modulator to the isolation stage and its 680 ohm grid leak resistor form a resistance-capacitance high pass-filter having a cut-off frequency of about 5 megacycles. This filter is necessary to prevent the 2.5-kilocycle signal at the modulator from finding its way into the synchronous detector where it would result in a false output signal. The 2.5-kilocycle voltage at the modulator output is caused by modulator plate current flowing through the power-supply impedance. The output of the isolation amplifier is coupled into the synchronous detector through another resistance-capacitance high pass filter to reduce the 2.5-kilocycle signal to a negligeble level.

<u>The synchronous detector</u>.--The synchronous detector is designed to give linear detection. It receives a "chopped" signal from the signal channel and has an output voltage at the chopping frequency of 2.5 kilocycles, which is directly proportional to the input signal voltage.

A reference signal is applied to the suppressor grids of the 6AS6 tubes used in this detector. Since this signal is of large amplitude, it is rectified by the suppressor grid and any modulation that may be present tends to appear in the output. This effect is most noticeable at very low input levels when noise may appear on the reference signal. To reduce this effect, two tubes are used as a balanced detector and the noise output which appears is the difference in the noise output of the two tubes. A considerable increase in sensitivity can be obtained in this manner. The circuit diagram of this balanced synchronous detector is shown in Fig. 6.

The operation of the synchronous detector will be shown, first, when the suppressor grid voltage is of large amplitude

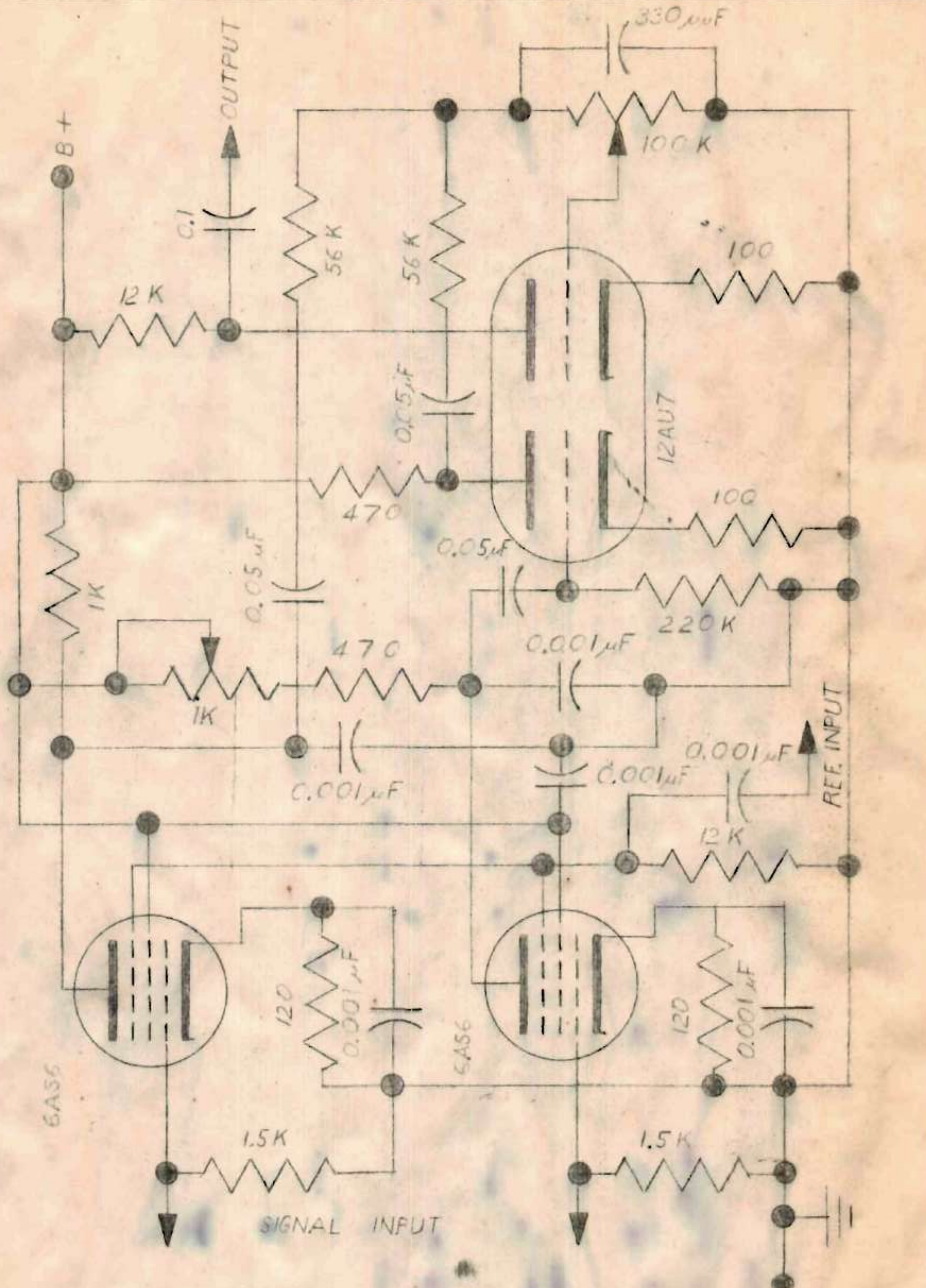


Fig. 6. Balanced Synchronous Detector and Phase Inverter

and square waveform and second when the suppressor grid voltage is sinusoidal and of low amplitude.

For the case of a square wave of large amplitude on the suppressor grid, and of the same frequency as a sinusoidal voltage applied to the control grid with an angular difference of ϕ degrees between the grid voltage and the fundamental component of the square wave, the operation of the synchronous detector may be expressed in the following manner.

The input signal voltage is amplitude modulated by the function, $1 + E_m(t)$. Let ω be the carrier frequency and the input signal to the synchronous detector may be expressed as

$$e(t) = \left[1 + E_m(t)\right]\left[\cos(\omega t + \phi)\right]$$

The transconductance, G_m, of the tube is modulated by a square wave such that

$$g_m(t) = G_m/2 + \sum_{n=1}^{n=\infty} (2G_m/n\pi)(-1)^{\frac{1}{2}(n-1)} \cos n\omega t$$

$$g_m(t) = G_m/2 + (2G_m/\pi)\cos\omega t - (2G_m/3\pi)\cos 3\omega t + (2G_m/5\pi)\cos 5\omega t \;.\;.\;.$$

The plate current, $i_p(t)$, of the detector may be expressed as

$$i_p(t) = \left[1 + E_m(t)\right]\left[\cos(\omega t + \phi)(G_m/2 + (2G_m/\pi)\cos\omega t - (2G_m/3\pi)\cos 3\omega t \;.\;.\;.)\right.$$

$$i_p(t) = G_m/2\left[1 + E_m(t)\right]\left[\cos(\omega t + \phi)\right] + \left[2G_m/\pi\right]\left[1 + E_m(t)\right]\left[(\tfrac{1}{2} + \tfrac{1}{2}\cos 2\omega t)\cos\phi\right.$$

$$+ \tfrac{1}{2} (\sin 2\omega t) \sin \phi\Big] - \Big[2G_m/3\pi\Big]\Big[1 + E_m(t)\Big]$$
$$\Big[(\tfrac{1}{2} \cos 2\omega t + \tfrac{1}{2} \cos 4\omega t) \cos \phi + (\tfrac{1}{2} \sin 4\omega t$$
$$- \tfrac{1}{2} \sin 2\omega t) \sin \phi \ . \ . \ .\Big]$$

It can be seen that the plate current of the tube contains a direct current component, plus the carrier frequency and its even harmonics. All are modulated by the same envelope as the input signal. If the carrier frequency and its harmonics are removed by a filter, the output current, $i(t)$, may be expressed as

$$i(t) = \Big[G_m/\pi\Big]\Big[1 + E_m(t)\Big] \cos \phi$$

The waveform of this output current is the same as that of the modulation envelope of the input signal.

In the second case, consider the operation of a synchronous detector in which a sine wave of low amplitude is applied to the suppressor grid. Let ω be the carrier frequency and the input signal be modulated by the function, $1 + E_m(t)$, such that

$$e(t) = \Big[1 + E_m(t)\Big] \cos (\omega t + \phi)$$

The voltage applied to the suppressor grid is

$$e_{g3}(t) = A \cos \omega t$$

where A is a constant. For a linear relation between the transconductance and the suppressor grid voltage, the control-grid-to-plate transconductance may be expressed in the form

$$g_m(t) = B\,(1 + A \cos \omega t)$$

where B is the transconductance of the tube with no signal voltage on the suppressor grid. The plate current can now be written as

$$i_p = \left[1 + E_m(t)\right]\left[\cos(\omega t + \phi)\right](B)(1 + A\cos\omega t)$$

$$i_p = \left[1 + E_m(t)\right]\left[B\cos(\omega t + \phi) + (\tfrac{1}{2}AB\cos\phi)(1 + \cos 2\omega t) + (\tfrac{1}{2}AB\sin\phi)(\sin 2\omega t)\right]$$

If all functions of ωt are removed with a filter, the output current becomes

$$i(t) = \left[1 + E_m(t)\right]\tfrac{1}{2}AB\cos\phi$$

As before the output current contains the same envelope as the input signal.

A third case to be considered is the effect of a square-law ratio of transconductance to suppressor grid voltage. Let ω be the carrier frequency and let the signal voltage be modulated by the function, $1 + E_m(t)$, such that the signal applied to the synchronous detector may be written

$$e(t) = \left[1 + E_m(t)\right]\left[\cos(\omega t + \phi)\right]$$

The voltage applied to the suppressor grid is

$$e_{g3}(t) = A \cos \omega t$$

where A is the amplitude of the voltage.

The transconductance is

$$g_m(t) = B\,(1 + A \cos \omega t)^2$$

where B is the transconductance of the tube with no voltage applied to the suppressor grid.

For these conditions the plate current can be expressed as follows:

$$i_p = \left[1 + E_m(t)\right]\left[\cos(\omega t + \phi)\right](B)(1 + A \cos \omega t)^2$$

$$i_p = \left[1 + E_m(t)\right]\left[\cos(\omega t + \phi)\right](B + 2AB \cos \omega t + BA^2 \cos^2 \omega t)$$

$$\begin{aligned} i_p = \left[1 + E_m(t)\right]\Big[&AB \cos \phi + B \cos(\omega t + \phi) \\ &+ \tfrac{1}{4}A^2B \left[(3 \cos \omega t) \cos \phi - (\sin \omega t) \sin \phi\right] \\ &+ AB \left[(\cos 2\omega t) \cos \phi + (\sin 2\omega t) \sin \phi\right] \\ &+ \tfrac{1}{4}A^2B \left[(\cos 3\omega t) \cos \phi - (\sin 3\omega t) \sin \phi\right]\Big] \end{aligned}$$

If the carrier frequency components and their harmonics are removed by a filter, the output current may be expressed as

$$i(t) = \left[1 + E_m(t)\right] AB \cos \phi$$

A linear relation between input and output is again obtained. It is not, however, the same output as was obtained for a linear relation between suppressor grid voltage and transconductance. This indicates that if the tube characteristics vary during operation, the output may be non-linear.

The sensitivity obtainable will be reduced by the presence of noise from the reference channel. The reference signal is rectified by the suppressor grid of the 6AS6 tube and any noise modulation of the reference signal will appear in the output. Since the amplitude of the reference signal is much greater than the amplitude of the signal being measured, a small amount of noise in the reference signal may obliterate the desired signal. To reduce this effect a push-pull detector is used. This tends to cancel any output signal that is a result of rectification, or square law response of the control or suppressor grids.

Effect of feedback.--Any feedback from the synchronous detector, which has a high-level reference signal, back to the low-level signal channel will result in very non-linear response. The output will be proportional to the vector sum of the signal and feedback voltages and may also be reduced if any phase shift is caused by the feedback. It is possible for the output to go to zero at a high signal level and then reverse polarity and increase in amplitude as the signal level is decreased, until the signal approaches the noise level. If the feedback voltage is in phase with the signal voltage, the output will approach some fixed level as the input signal decreases. This output level will be maintained until the signal approaches the noise level of the circuit.

To produce the effects noted above, the feedback must be to a part of the signal channel preceeding the modulator. If the feedback is to a part of the circuit following the modulator, the effect will be to produce a direct current component in the output. This feedback will not cause non-linearity unless it is of sufficient amplitude to overload the signal channel amplifier.

CHAPTER III

EXPERIMENTAL RESULTS

Description of the receiver.--The receiver is designed to be used as a fixed-frequency detector or as the intermediate-frequency amplifier of a very high frequency or microwave receiver. The input signal feeds two channels. One, the reference channel, has high gain and gives constant output. The other, the signal channel, has a linear response, and contains a modulator, which chops the signal at a frequency of 2,500 cycles. It also contains a delay line which gives it a phase-shift-versus-frequency characteristic which closely approaches that of the reference channel.

The outputs of these two channels are combined in a balanced synchronous detector. The output of the reference channel is fed to the suppressor grids of the tubes in parallel. The modulated push-pull output of the signal channel is fed into the control grids. The detected output from the synchronous detector is push-pull, while any output produced by rectification or non-linear response by the control grids or suppressor grids gives in-phase output from the two detector tubes. The output of one detector tube is reversed by means of a phase inverter and combined with the output of the other detector tube. Thus, signal components of the detector's output add, while components, due to non-linear response, cancel. A gain control is incorporated in the phase inverter to make it possible to obtain more complete cancellation of the in-phase components by compensating for differences in tubes and component values. This results in a considerable increase in sensitivity as will be shown later in this chapter.

The reference channel is shielded from the signal channel, since stray coupling between them may result in non-linear response.

An automatic frequency control circuit is included. It is designed for use with an external microwave converter that uses a reflex-klystron tube as its local oscillator. Any klystron tube which operates with less than 4,000 volts on its repeller may be used.

Plate voltage is obtained from a regulated power supply delivering about 120 volts direct current. Regulation is used to keep the power-supply hum level low and to prevent plate voltage variations which might change the gain of the unit. A complete diagram of the receiver is shown in Fig. 7. Component values are not shown, but they may be obtained from the circuit diagrams given in Chapter II.

<u>Difficulties encountered in tests on the receiver</u>.--Feedback of the reference signal into the signal channel will occur unless the two channels are very well shielded. The feedback

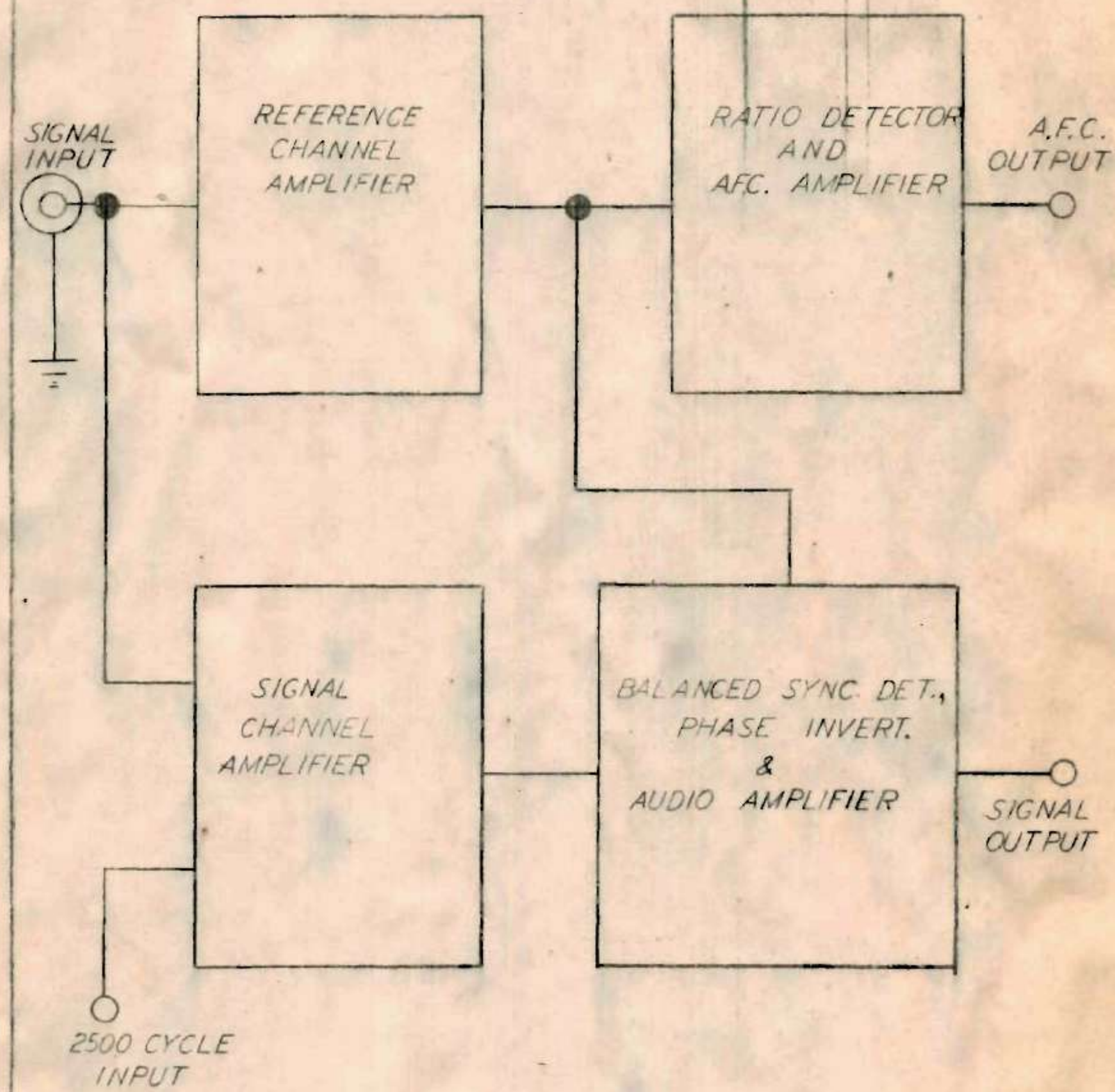


Fig. 7. Synchronous Detection Receiver

voltage adds to the signal voltage and may either increase or decrease the output, depending on the phase of the feedback voltage. A typical result of this feedback is for the output to decrease non-linearly to zero and then build up to some constant amplitude, but of opposite phase, and maintain this output until the signal level approaches the noise level of the circuit. Even with shielding, it was found advisable to use low impedance circuits to reduce stray pickup. The difference of more than 100 db in the signal level of the two channels when the input-signal level is low makes it very difficult to completely eliminate the feedback.

Some difficulty was encountered with 2,500 cycle voltage from the modulator finding its way into the synchronous detector and resulting in an output with no input signal, as well as affecting the linearity of the output. As explained earlier, the coupling circuits between the modulator and the synchronous detector are designed to be high-pass filters with cut-off frequencies of about five megacycles. This reduces feed-through of 2,500-cycle voltage to a negligible level.

Differences in phase shift of the signal and reference channels will cause the output to have wide variations in amplitude with changes in carrier frequency. The output amplitude-versus-frequency resembles a damped-cosine wave, if the phase shifts of the reference and signal channels do not vary at the same rate. The response of this unit with and without the phase-correcting delay line is shown in Fig. 8.

<u>Results of the tests</u>.--The response of this synchronous detector is quite linear over a dynamic range of 80 db. Fig. 9 shows the range and linearity of the detector and the advantage of the balanced detector. An increase in sensitivity of about 13 db was obtained by balancing the detector. The signal source was a Hewlett-Packard type 608 signal generator. Its attenuator which has a tolerance of plus or minus one db was used in adjusting the input signal level. The output of the synchronous detector was attenuated to a constant level with a General Radio step attenuator. The output was monitored with a Tektronix oscilloscope and pre-amplifier.

The minimum signal which could be detected by the balanced synchronous detector was about -93 dbm. The unit operates best if the input signal level stays between -5 dbm and -85 dbm. Below -85 dbm, there is a noticeable amount of noise in the output. At levels greater than -5 dbm, the signal channel amplifier overloads. The difference between the output of the balanced synchronous detector and straight line representing linear response is 1 db at -5 dbm, 3 db at 0 dbm, and 9 db at 7 dbm. At -90 dbm the error is approximately 2 db. The presence of noise in the output at very low levels makes it difficult to measure the output accurately. The output voltage was observed on an oscilloscope, since a

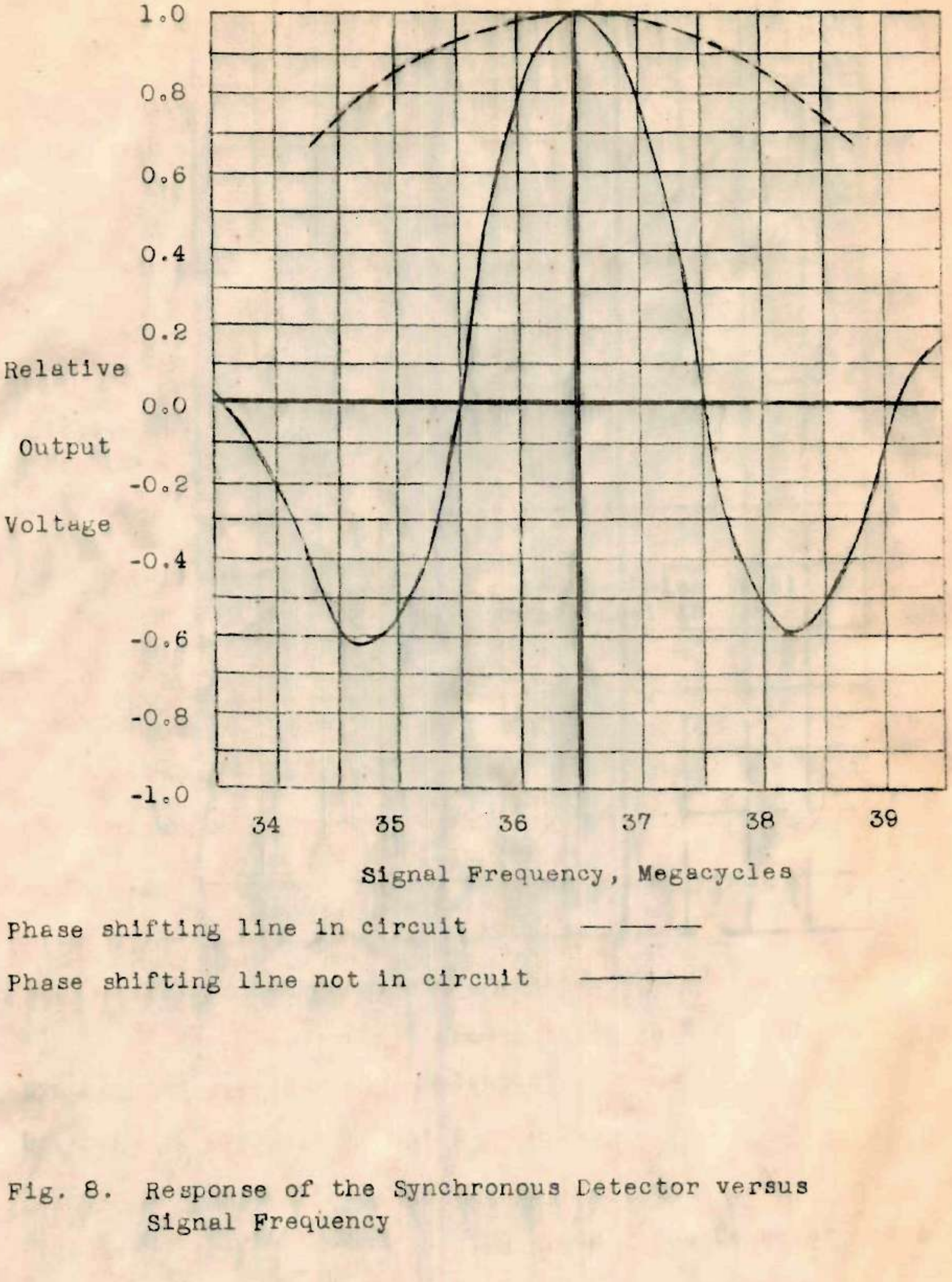
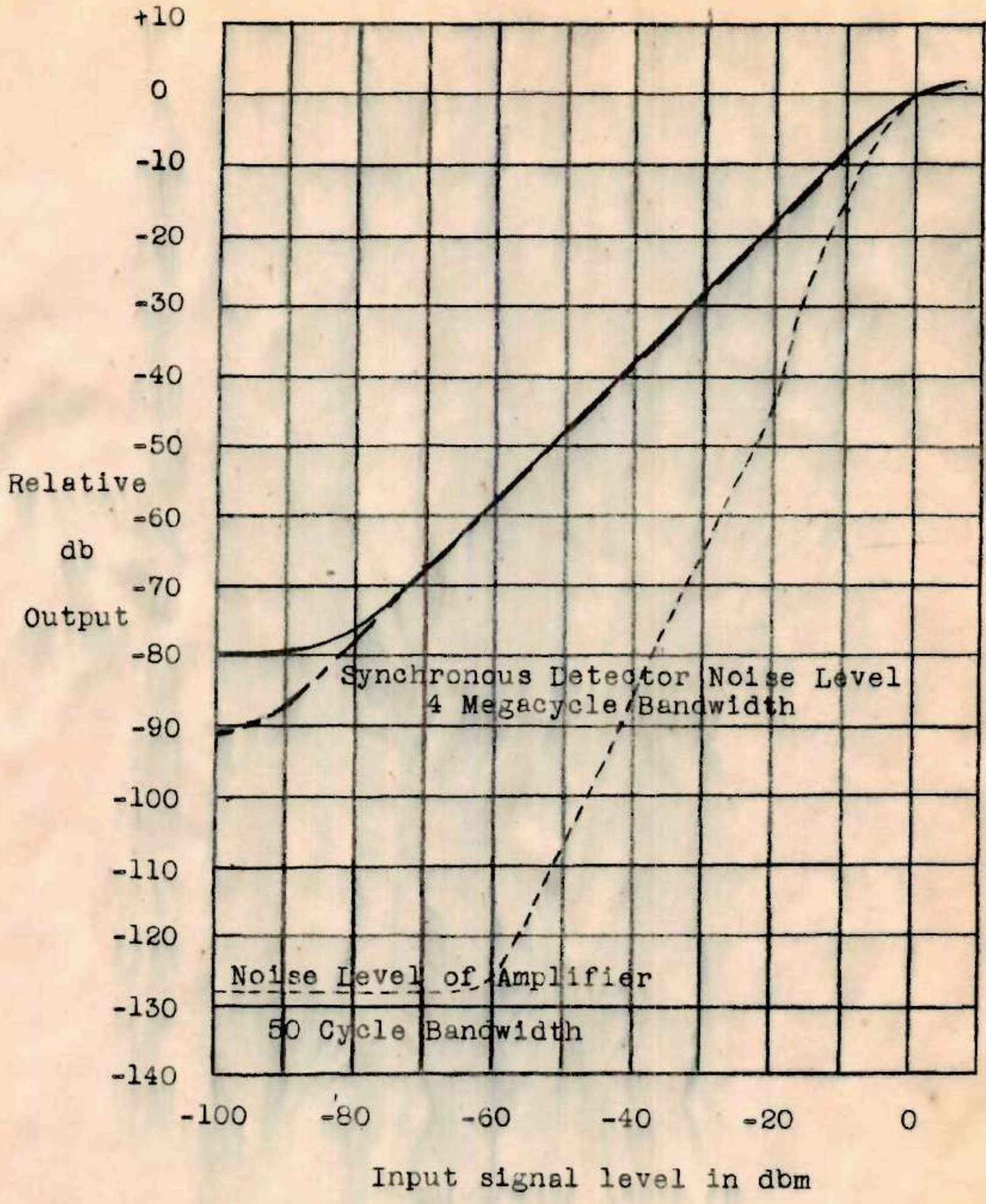


Fig. 8. Response of the Synchronous Detector versus Signal Frequency



Synchronous detector (not balanced) ———

Balanced synchronous detector — —

1N23 crystal diode at 6,300 megacycles -----

Fig. 9. Relative Response of the Synchronous Detector and a 1N23 Crystal Diode

vacuum tube voltmeter would tend to read the level of the noise peaks plus the signal.

For comparison, the response of a 1N23 crystal detector is also shown in Fig. 9. A carrier frequency of 6,300 megacycles, with 1,000-cycle square-wave modulation, was supplied by a Hewlett-Packard Model 618A signal generator. The rectified signal level was measured with a Hewlett-Packard Model 415A voltmeter. This meter contains an attenuator and a 1,000-cycle amplifier which has a bandwidth of 50 cycles. The very narrow band-pass of the amplifier made it possible to measure much lower output levels than could be measured for the synchronous detector, which has a bandwidth of four megacycles.

All three curves in Fig. 9 have been referred to the same level at 0 dbm in order that they may be compared easily.

The response-versus-frequency is shown in Fig. 8. The oscillating response obtained when the delay line is not in the circuit is very similar to the variation predicted by the calculations of Chapter II. The frequencies nearest the center of the band at which the output becomes zero are separated by two megacycles. The difference in the phase shift of the reference channel and the signal channel is therefore 90 degrees per megacycle. This is 9 degrees more than the calculations predicted and is probably caused by the Q of the tuned circuits being slightly higher than the values used in calculating the phase shift.

The response with the delay line in the circuit is a smooth curve which is approximately 3 db down from its peak when the intermediate frequency is 2.0 megacycles either side of center. This 3 db point may be due in part to the small remaining difference in the phase shifts of the two channels.

The response of the automatic frequency control unit is shown in Fig. 10. The response has been made very sharp in order to hold the carrier near the center frequency of the detector. This prevents error that would be caused by the drop in response near the edges of the detector's frequency band. The automatic frequency control voltage does not begin to decrease until the frequency is about four megacycles either side of the center frequency. This makes it improbable that the circuit would hold the carrier away from the center frequency by operating outside the peaks of the ratio detector's response where the slope of the voltage-versus-frequency curve reverses.

The two klystron tubes required for testing the automatic frequency control were not available. In order to show the improvement in frequency stability produced by this unit, a theoretical curve of the intermediate frequency as a function of the received carrier frequency is shown in Fig. 11. The local oscillator is assumed to be a klystron tube whose frequency changes one megacycle per volt change in repeller

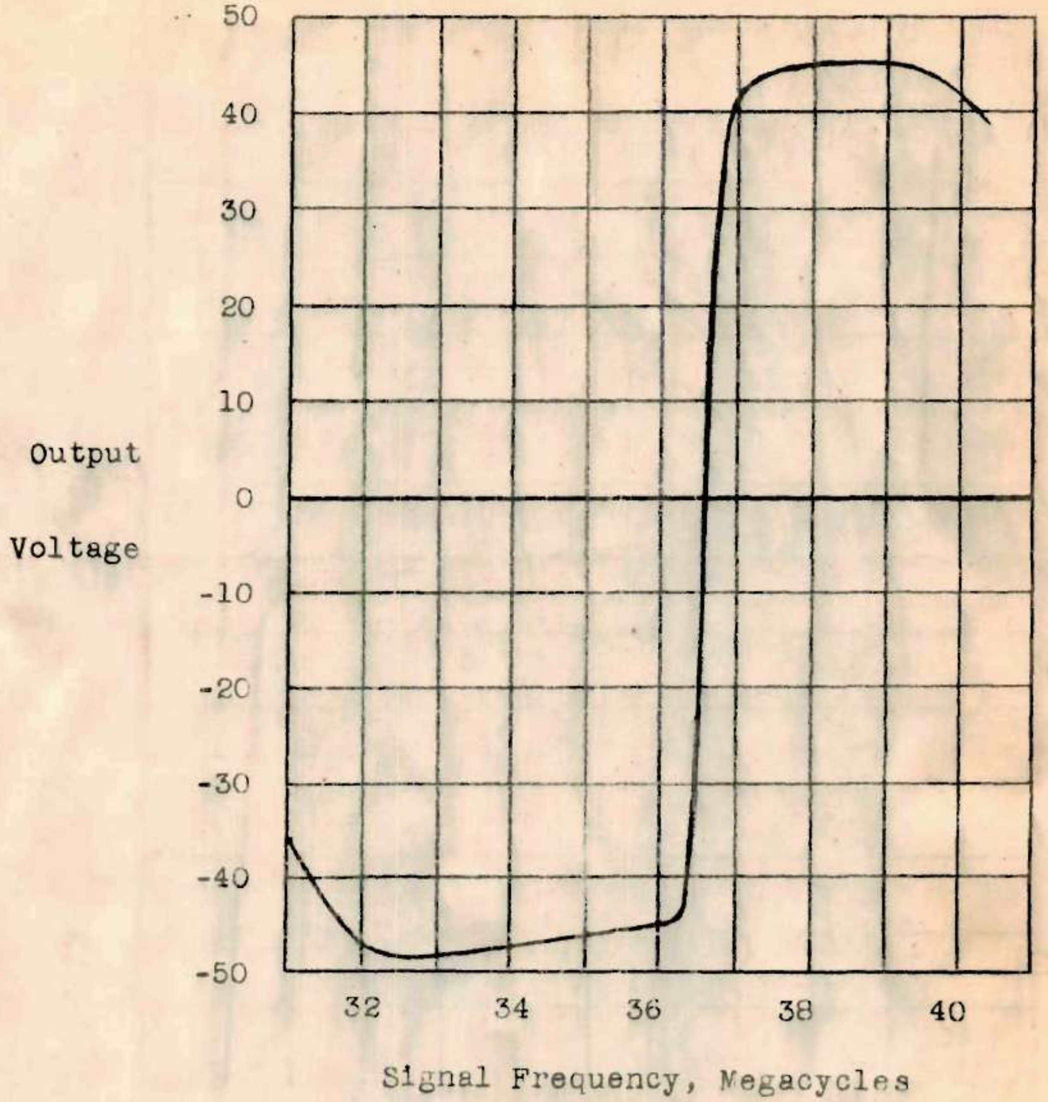


Fig. 10. Automatic Frequency Control Output Voltage as a Function of Signal Frequency

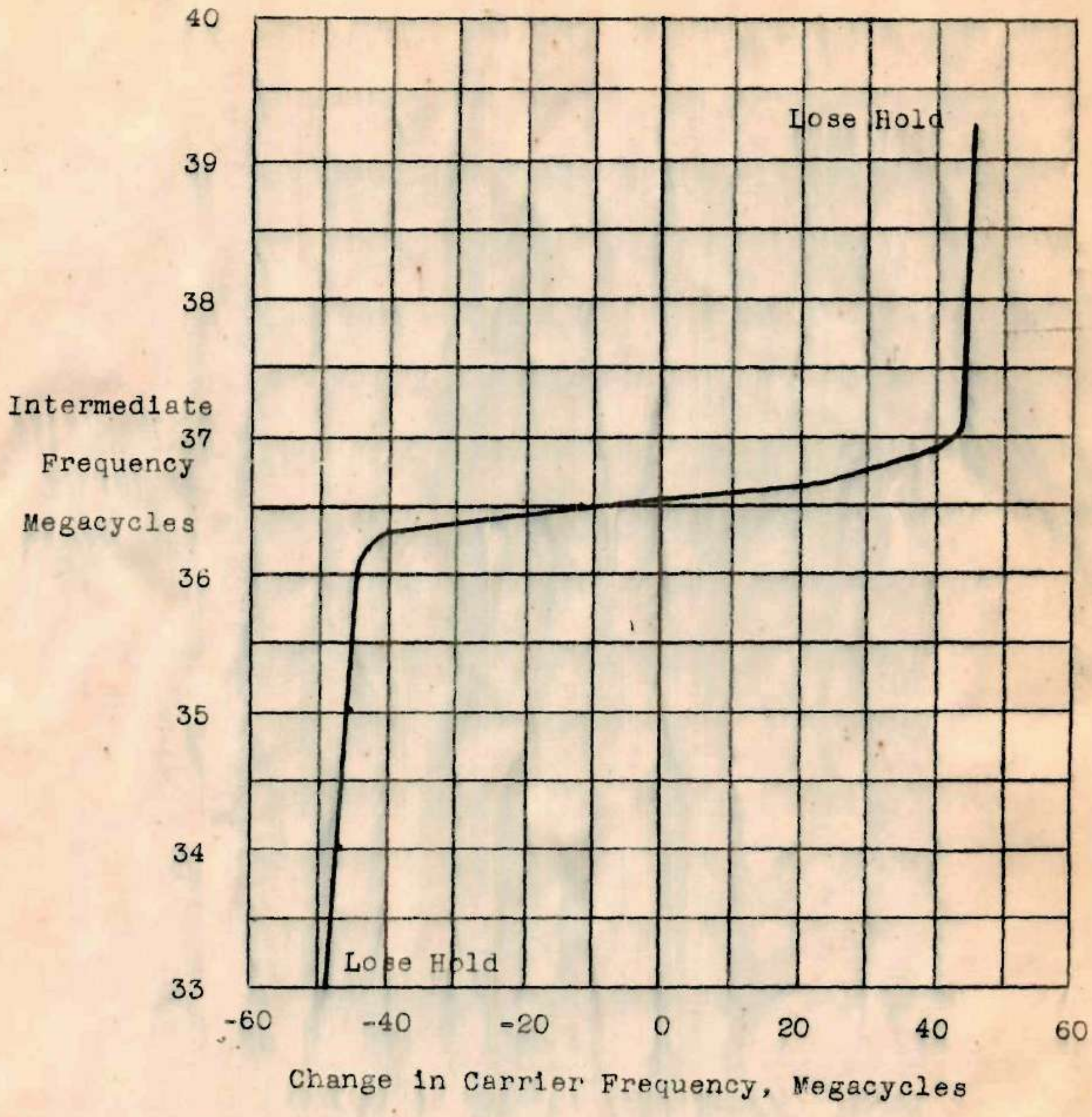


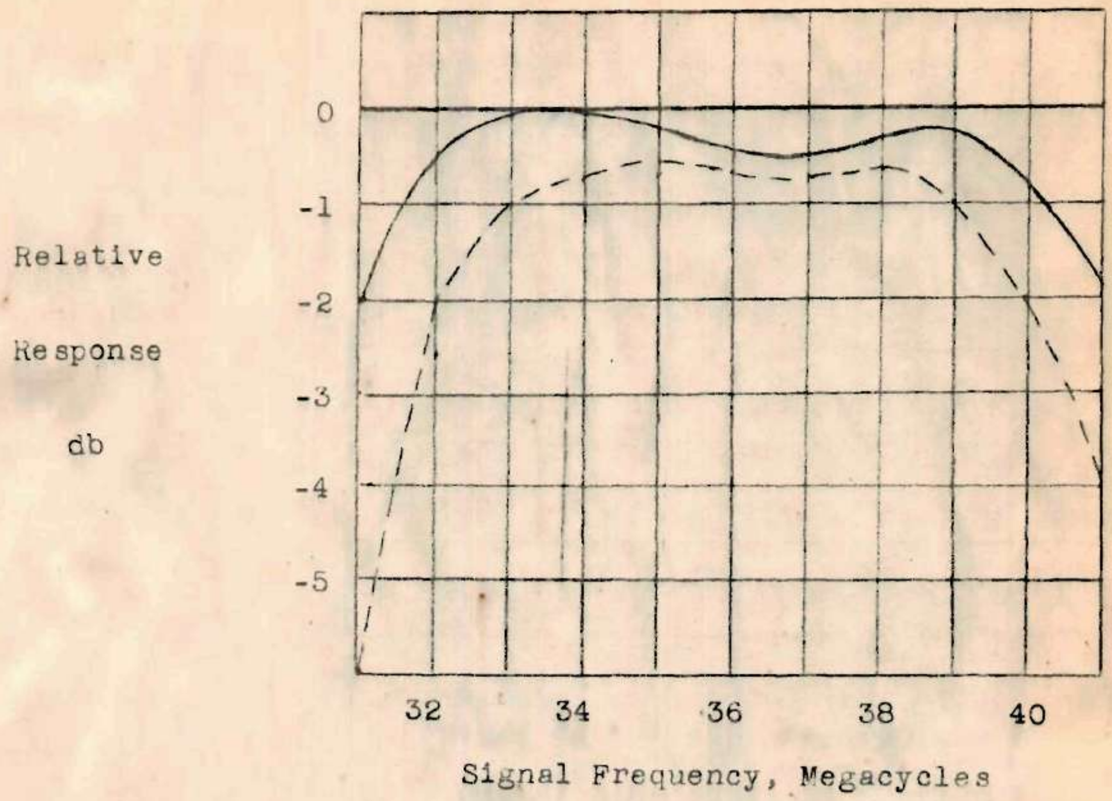
Fig. 11. Intermediate Frequency as a Function of the Received Carrier Frequency for a Converter whose Local-Oscillator Frequency changes 1.0 Megacycle per Volt of Automatic Frequency Control Voltage

voltage.

The response of the reference channel amplifier is shown in Fig. 12. Its response is relatively flat over the bandwidth of the detector. If the input signal level is low, the limiters will not remain saturated over as wide a band as when the signal level is high. This will cause the reference channel amplifier to have a greater bandwidth when the input signal level is high than when it is low. The output of the reference channel amplifier versus frequency for both high and low input signal levels is shown in Fig. 12.

The amplitude-versus-frequency response of the signal channel is shown in Fig. 13. It was checked at a high input level since the voltmeter which was available could not be used to measure less than 0.1 volt.

With no phase error, the output of the synchronous detector is proportional to the product of the output of the reference channel and the output of the signal channel. Each channel has a bandwidth greater than that required for the unit in order to prevent the bandwidth from becoming too narrow if a small amount of phase error were introduced by changing a tube or by a slight change in a component value.



Input-signal level of 0 dbm ————

Input-signal level of -60 dbm – – – – –

Fig. 12. Response of the Reference Channel Amplifier as a Function of the Signal Frequency

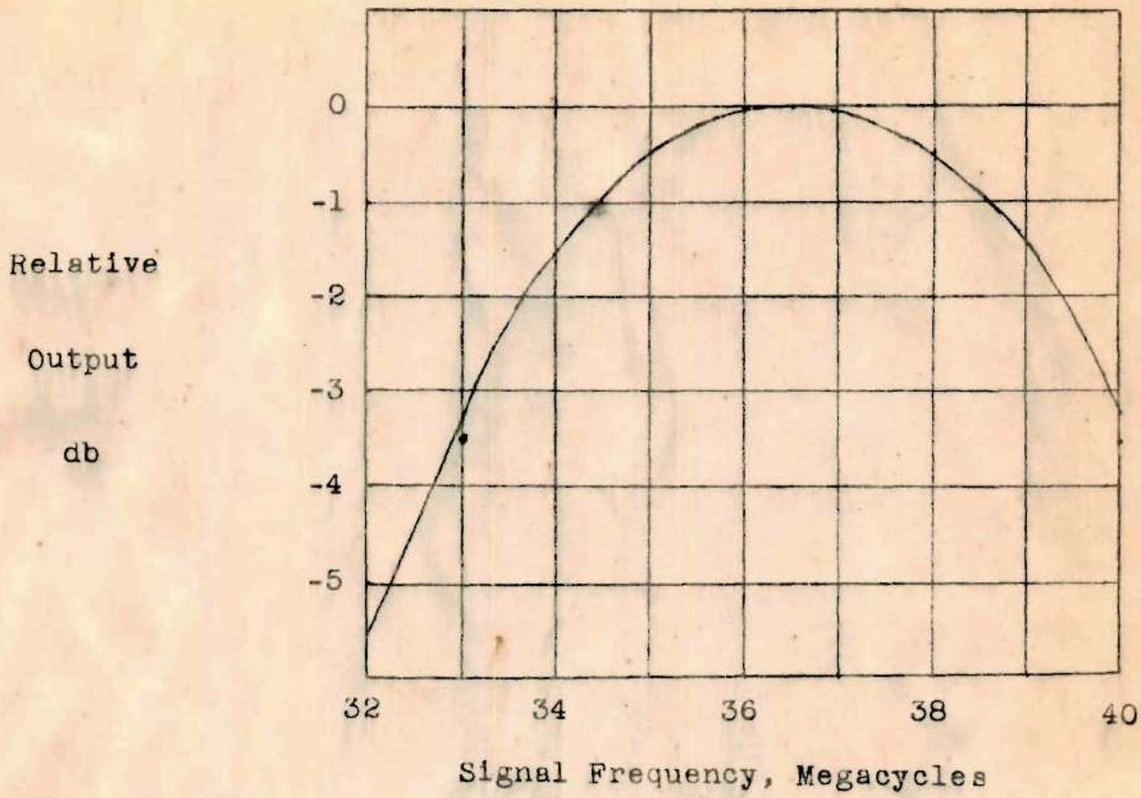


Fig. 13. Signal-Channel Output as a Function of Signal Frequency

CHAPTER IV

CONCLUSIONS

Logarithmic recorders with a range of 80 db are available. Low signal frequencies, such as 1,000 cycles, are normally used. If it is desirable to record a high-frequency signal, some means of frequency conversion, or detection of modulation on the signal must be used. Most detection systems have a square-law response which would limit the recording range to 40 db. Conversion of the high-frequency signal to a frequency of about 1,000 cycles could be done, although it might be very difficult to obtain the necessary frequency stability of the local oscillator if the carrier being measured had a frequency of 10,000 megacycles.

The synchronous detector provides an easy method of obtaining the full range of a recorder, or other measuring device, at a very high frequency.

Although the unit described in this thesis has a bandwidth of about four megacycles, a similar type of unit could be constructed for almost any bandwidth which could be obtained with the reference channel.

If very high sensitivity or very wide range is needed from a unit of this type, a narrow bandwidth should be used.

The noise level in the reference channel limits the sensitivity of this unit. If it could be lowered, greater dynamic range could also be obtained. About 80 db of linear range was obtained with the synchronous detector described in this paper. The low-level end of the range was limited by noise voltage from the reference channel, while the high-level end was limited by overload of the amplifier tubes in the signal channel.

The increase in range and sensitivity over a square-law detector makes it possible to record the same dynamic range with a much lower power level if desired. In tests with this detector, an improvement of about 30 db in sensitivity with respect to a 1N23 crystal was obtained.

The high sensitivity of the synchronous detector could be used to advantage in any microwave measurements in which a magnetron tube is ordinarily needed to provide the necessary power. It would make it possible to use a small klystron tube in place of the magnetron tube, although an additional klystron tube would be required to provide the local oscillator signal for the receiver.

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